the noise contributed by the IF amplifier. Fig. 4 shows how the measured noise figure varied with respect to LO power, with the LO frequency as a parameter. The 1.5-GHz IF contributed 5 dB to the measurement, and the dc bias voltage was varied to maintain a constant current of 1.5 mA through the diode. The measured noise figure was 9-10 dB for LO frequencies in the 60-61-GHz range for drive levels of 0.8-1.8 mW.

Fig. 4 shows that an overall noise figure of 6 dB is feasible in future models with superior FET IF amplifiers. Owing to the lower losses of high-quality microstrip, its freedom from tight tolerances and compatibility with hybrid devices, this new transmission line is well suited to a wide variety of integrated circuits throughout the millimeter bands.

ACKNOWLEDGMENT


REFERENCES


Simplified 12-GHz Low-Noise Converter with Mounted Planar Circuit in Waveguide

YOSHIHIRO KONISHI, SENIOR MEMBER, IEEE, KATSUAKI UBAKADA, NORIHITO YAZAWA, NOLIO HOSHINO, AND TADASHI TAKARASHI

Abstract—A 12-GHz low-noise converter consisting of a planar circuit mounted in a waveguide was described in this paper. It consists of a planar circuit mounted in a waveguide and is suitable for low-cost mass production.

Millimeter components using integrated circuits mounted in a waveguide were also developed by Meier in 1972 [2], where unloaded Q of this transmission line takes a value less than 900 at X band [3].

A new type of filter, that is, a planar circuit mounted in waveguide, that we have developed has an unloaded Q factor of 2500-3000 at X band. This new type of filter is used in our 12-GHz converter to achieve low-noise performance.

The 12-GHz converter described in this paper has a mixer conversion loss of 3.5 dB and a total noise figure of 4.5 dB, including the contribution of an intermediate frequency amplifier with a noise figure of 2.0 dB at 420 MHz.

II. 12-GHz LOW-NOISE CONVERTER WITH MOUNTED PLANAR CIRCUIT IN WAVEGUIDE

A high-sensitivity and low-cost converter was required that could be constructed simply and be mass produced. The construction of the circuit we developed is such that every necessary circuit element is arranged on a metal sheet merely by pressing or etching, and the metal sheet is inserted into a waveguide.

We will describe the results of the experiments that were carried out on our converter with a mounted planar circuit.

Fig. 1 shows the planar circuit pattern from left to right, a signal frequency band-pass filter, a Schottky barrier diode mount, a local oscillator frequency band-pass filter, and a Gunn diode mount for the local oscillator. A 0.3-0.5-mm-thick copper sheet is favorable for this pattern. When etching is used in forming the pattern, a 0.3-mm copper sheet is utilized for dimensional precision. When pressing is used, with a 0.3-mm-thick metal sheet, dimensions are maintained within 20 μm.

Manuscript received July 31, 1973; revised December 4, 1973.

The authors are with the Technical Research Laboratories, Japan Broadcasting Corporation, Tokyo, Japan.
As $Z$ is varied by changing the post width, impedance matching becomes possible for a low-impedance diode, too.

The curves in Fig. 3(b)–(d) show the impedance viewed from the diode terminal when the post width $W$ and the height of the ridge taper $A$ are changed. Fig. 3(b) shows the impedance characteristics for various post widths $W$ in the case of $\lambda/4$ post length. An approximate change is made as was described previously. Fig. 3(c) shows the impedance when $W$ of the flat post remains constant at 4 mm and a change is made in the height of the taper. Fig. 3(d) shows the impedance when the width $W$ is changed only by the flat post, the tip of which is shortened. In Fig. 3(d), in comparison with Fig. 3(b), there is an increase in reactance and the bandwidth proves to be narrower. According to chart $a$, we can obtain arbitrary impedance in wide bands by selecting appropriate $W$ and $A$.

C. Consideration of Image Impedance

It is necessary to consider the image impedance in reducing the conversion loss [4]. At the image frequency, the signal frequency and the local oscillator frequency bandpass filters, which are arranged at both sides of the diode mount, produce a short circuit at an inside point a little off the end surface of the filter. Accordingly, the distance between the two filters varies the image frequency impedance that is presented to the diode. In general, where the distance between the filters is selected equal to $m\lambda/2$ at the image frequency, the image impedance is an open circuit. As the position of the signal frequency and the local oscillator frequency bandpass filters are adjusted to optimize the signal frequency and local oscillator frequency impedances, it is not possible to provide the exact image frequency impedance for minimum conversion loss.

For the converter described in this short paper, the image impedance remains a little off the open condition and capacitive.

D. Local Oscillating Circuits

The local oscillating source used a Gunn diode, and the construction described in Section II-B is used as a mount.

III. RESULTS OF OUR EXPERIMENT ON PROPOSED CONVERTER

We will describe the results of our measurements on the experimental converter with the metal sheet on which every foregoing pattern is formed, as shown in Fig. 1.

Fig. 4 shows the characteristics of the converter we have developed. The local oscillator drive is 8 mW. The signal frequency is 12 GHz and the intermediate frequency is 420 MHz. We were successful in obtaining a minimum real conversion loss of 3.2 dB. An intermediate amplifier with approximately a 2.0-dB noise figure was connected and a total noise figure 4.5 dB was obtained. This value coincides with the result of the noise figure equation according to the Appendix.

APPENDIX

Overall receiver noise figure $F_r$ can be expressed as

$$F_r = LsL_m(L + F_m - 1)$$  \quad (1)

where

- $L_r$ RF circuitry loss;
- $L_m$ mixer conversion loss;
- $F_m$ noise figure of IF amplifier;
- $T_m$ mixer noise temperature.

In the case of the image resistive load, the noise figure $F$ of the mixer diode is given by [6]

$$F = 1 - n + nL$$  \quad (2)

where

- $n = \frac{e}{2kT_a}$ equals noise ratio;
- $e$ charge on an electron;
- $k$ Boltzmann’s constant;
- $T_a$ temperature;
- $\alpha$ parameter of Schottky barrier diode (see [6]).

In the case of $\alpha = e/(kT_a)$, the noise ratio is equal to 0.5.
Fig. 3. Relationships between diode mount and impedance. (a) Diode mount. (b)-(d) Impedance or admittance coordinates.

Fig. 4. Characteristics of 12-GHz converter with mounted planar circuit in waveguide.
for, the noise figure $F$ is given by

$$ F = \frac{1}{2} (1 + L). $$

(3)

This is the same formula as in [3]. Also, the equation relating $t$, $L$, and $F$ is

$$ t = F/L. $$

(4)

Substituting formula (3) into (4) yields

$$ t = \frac{1}{L} (1 - n) + n. $$

Measured values for the mixer were

$$ L = L = 2.0 \text{ (3.2 dB)} $$
$$ F = 1.585 \text{ (2.0 dB)} $$
$$ n = 32 \text{ (4.57 dB)} $$

Calculated values were

$$ n = 0.501 $$
$$ t \approx \frac{1}{L} (1 - n) + n = 0.7555 $$
$$ F = 2.86 \text{ (4.57 dB)} $$

REFERENCES


Efficient Numerical Computation of the Frequency Response of Cables Illuminated by an Electromagnetic Field

CLAYTON R. PAUL, MEMBER, IEEE

Abstract—Computationally efficient numerical methods for determining the frequency response of uniform transmission lines consisting of a large number of mutually coupled conductors in homogeneous and inhomogeneous media, and illuminated by an electromagnetic (EM) field are presented.

I. INTRODUCTION

Conductors connecting electronic subsystems on aircraft, missiles, and ground electronic systems are generally grouped into large, closely coupled cable bundles and it is not uncommon to find bundles of over 100 conductors on modern avionic systems (for example, F-4, F-111, and F-15 aircraft) [3]. Determination of the frequency response of these large cable bundles illuminated by high-power radars as well as an electromagnetic pulse (EMP) from nuclear detonations is becoming of increasing importance [1]. Computation of the frequency response of these large bundles illuminated through an aperture such as a landing gear door can be quite costly for only one frequency. However, it is generally necessary to determine the response for many frequencies so the consideration here is to minimize the per-frequency computation time for bundles consisting of a large number of conductors. Flat pack and woven flat cables are being used more frequently to connect electronic subsystems and it is not uncommon to find over 35 mutually coupled conductors in these types of cables [7].

Taylor et al. [2] considered the problem of two conductors illuminated by a nonuniform electromagnetic (EM) field. For two conductors, the per-frequency computation times are practically minimal. For larger numbers of conductors, we encounter per-frequency computation times which are functions of $n$ for an $(n+1)$ conductor line so that reduction of the per-frequency computation times becomes an important concern for large numbers of coupled conductors and many computed frequencies.

We will cast the equations to be solved at each frequency into particularly efficient forms as well as introduce computational procedures peculiar to these forms which allow an efficient solution. Perhaps many of the results here will be considered fairly straightforward to obtain but our purposes will be to unify the particular formulations and also point out some perhaps not so obvious techniques for reducing computation times.

Consider an $(n+1)$ conductor uniform transmission line consisting of $(n+1)$ parallel lossless conductors of length $L$ embedded in a lossless dispersive medium with the $(n+1)$st conductor designated as the reference conductor (usually a ground plane or overall shield). The transmission line is described for the TEM mode by the following 2n strongly coupled complex differential equations [3], [5]:

$$ \left[ \begin{array}{c} V(z) \\ I(z) \end{array} \right] = \begin{bmatrix} 0 & -i\omega & L & 0 \\ -i\omega & 0 & 0 & L \end{bmatrix} \left[ \begin{array}{c} V(z) \\ I(z) \end{array} \right] $$

(1)

where $V(z) = (d/dz) V(z)$ and $I(z)$ is the $i \times j$ zero matrix. The distance along the conductor structure and parallel to it is denoted by $z$; the complex currents $I(z)$ are directed in the direction of increasing $z$ and the $i$th elements of the $n \times 1$ vectors $V(z), V'(z)$, and $I(z), I'(z)$ are the complex potentials (with respect to the reference conductor) and currents, respectively, associated with the $i$th conductor, $i = 1, \cdots, n$. The parameter $\omega$ is the radian frequency of excitation under consideration and $n \times n$ real symmetric constant matrices $\mathbf{L}$ and $\mathbf{C}$ are the per unit length inductance and capacitance matrices, respectively [3], [5].

The boundary conditions at the ends of the transmission line are in the form of $n$ ports and are characterized by "generalized Thewlein equivalences" as

$$ V(0) = \mathbf{E}_0 - \mathbf{R}_0 I(0) $$

(2a)

$$ V(L) = \mathbf{E}_0 + \mathbf{R}_0 I(L) $$

(2b)

where $\mathbf{E}_0$ and $\mathbf{R}_0$ are $n \times 1$ complex vectors of the equivalent open circuit port excitations and $\mathbf{R}_0$ and $\mathbf{R}_0$ are $n \times n$ real symmetric hyperdeterminant (and therefore positive definite) matrices representing passive termination networks.

Initially, we must solve (1) and then we must incorporate the boundary conditions of (2). Differentiating the second equation in (1) with respect to $z$ and substituting the first we obtain

$$ I(z) = -\omega^2 \mathbf{C} I(z) $$

(3)

where $\mathbf{R}_0$ is a $n \times n$ diagonal matrix with real positive and nonzero scalars $\gamma_i^2$ on the diagonal, i.e., $[\mathbf{R}_0]_{ii} = \gamma_i^2$ and $[\mathbf{R}_0]_{ij} = 0$ for

$$ T^{-1} \mathbf{C} \mathbf{L} T = \mathbf{I}^2 $$

(5)